# TITLE OF THE INVENTION

## HIGH TEMPERATURE SUPERCONDUCTING TUNABLE FILTER

## CROSS-REFERENCE TO RELATED APPLICATIONS

5      [0001]      This is a continuation application of Appl. Ser. No. 10/023,575, filed on

Dec., 17, 2001, and issuing as U.S. Pat. No. 6,662,029 on Dec. 9, 2003, which is a continuation

application of Appl. Ser. No. 09/268,786, filed on March 16, 1999, and issued as U.S. Pat. No.

6,347,237 on Feb. 12, 2002, the disclosures of which are hereby incorporated by reference.

10      ## FIELD OF THE INVENTION

[0002]      This invention relates to a high temperature superconductor (HTS) tunable

filter.  More particularly, this invention relates to an HTS filter tunable by variation of a

microelectromechanical capacitor.

15      ## BACKGROUND OF THE INVENTION

[0003]      The need for a high-quality factor (Q), low insertion loss tunable filter

pervades a wide range of microwave and RF applications, in both the military, *e.g.*, RADAR,

communications and electronic intelligence (ELINT), and the commercial fields such as in various

communications applications, including cellular.  Placing a sharply defined bandpass filter directly

20      at the receiver antenna input will often eliminate various adverse effects resulting from strong

interfering signals at frequencies near the desired signal frequency in such applications.  Because

of the location of the filter at the receiver antenna input, the insertion loss must be very low to not

degrade the noise figure. In most filter technologies, achieving a low insertion loss requires a

corresponding compromise in filter steepness or selectivity. In the present invention, the

extremely low loss property of high-temperature superconductor (HTS) filter elements provides an

5　attractive solution, achieving a very low insertion loss yet simultaneously allowing a high

selectivity/steepness bandpass definition.

[0004]　In many applications, particularly where frequency hopping is used, a

receiver filter must be tunable to either to select a desired frequency or to trap an interfering signal

frequency. Thus, the insertion of a linear tunable filter between the receiver antenna and the first

10　nonlinear element (typically a low-noise amplifier or mixer) in the receiver offers, providing that

the insertion loss is very low, substantial advantages in a wide range of RF and microwave

systems. For example, in RADAR systems, high amplitude interfering signals, either from

"friendly" nearby sources, or from jammers, can desensitize receivers or intermodulate with high-

amplitude clutter signal levels to give false target indications. In high-density signal

15　environments, RADAR warning systems frequently become completely unusable.

[0005]　Both lumped element and distributed element filters suffer from these and

other problems. For example, while distributed-element YIG-tuned filters have been used, the

high level of insertion loss (usually greater than 10 dB) of suitable YIG filters necessitates their

use on a "switch in when absolutely necessary" basis only, as the degradation to noise figure

20　would generally be unacceptable. Lumped element filters also suffer problems. For a lumped

element filter to be tunable, the filter requires either a tunable capacitor, or a tunable inductive

element. The vast majority of RF tunable lumped element filters have used varactor diodes. Such a design amounts to using a tunable capacitor because varactor diodes, through changing the reverse bias voltage, vary the depletion thickness and hence the PN junction capacitance. While varactors are simple and robust, they have limited quality factors (Q), and suffer from the problem

5      that the linear process that tunes them extends all of the way to the signal frequency, so that high-amplitude signals create, through the resulting nonlinearities, undesirable intermodulation products, etc. The same problems of poor Q and high-frequency nonlinearities are anticipated for "tunable materials" such as ferroelectrics.

        [0006]        Consider the case of a conventional varactor diode. In a varactor, the

10     motion of electrons accomplishes the tuning itself. As the reverse bias ($V_r$) on the junction of the varactor is changed, then in accordance with Poisson's Equation, the width of the PN junction depletion region changes which alters the capacitance of the junction ($C_j$). Because the tuning mechanism of varactors is electronic, the tuning speed is extremely fast. Unfortunately, this also leads to a serious associated disadvantage: limited dynamic range. Because the $C_j$ ($V_r$) relationship

15     is nearly instantaneous in response, extending to changes in $V_r$ at the signal frequency itself, and because the signal (frequently in a resonantly magnified form) appears as a component of the junction bias voltage, $V_r$, the signal itself parametrically modulates the junction capacitance. If the signal amplitude across the varactor is very small in comparison to the dc bias, the effect is not too serious. Unfortunately, for high signal amplitudes, this parametric modulation of the capacitance

20     can produce severe cross-modulation (IM) effects between different signals, as well as harmonic generation and other undesirable effects. While these signal-frequency varactor capacitance

variations are the basis of useful devices such as parametric amplifiers, subharmonic oscillators, frequency multipliers, and many other useful microwave circuits, in the signal paths of conventional receivers they are an anathema. This inherent intermodulation or dynamic range problem will presumably extend to "tunable materials," such as ferroelectrics or other materials in

5      which the change of dielectric constant ($\varepsilon_r$) with applied electric field (E) is exploited to tune a circuit. As long as the $\varepsilon_r$ (E) relationship applies out to the signal frequency, then the presence of the signal as a component of the E will lead to the same intermodulation problems that the varactors have.

[0007]      In addition to the intermodulation/dynamic range problems of varactors,

10      these conventional tuning devices also have serious limitations in Q, or tuning selectivity. Because the varactors operate by varying the depletion region width of a P-N junction, this means that at lower reverse biases (higher capacitances), there is a substantial amount of undepleted moderately-doped semiconductor material between the contacts and the junction that offers significant series resistance ($R_{ac}$) to ac current flow. Since the Q of a varactor of junction

15      capacitance $C_j$ and series resistance $R_{ac}$ at the signal frequency f is given by $Q = 1/(2 f C_j R_{ac})$, this means that the varactor Q values are limited, particularly at higher frequencies. For example, a typical commercial varactor might have $C_j = 2.35$ pF with $R_{ac} = 1.0 \; \Omega$ at $V_r = -4V$, or $C_j = 1.70$pF with $R_{ac} = 0.82 \; \Omega$ at $V_r = -10V$, corresponding to Q values at $f = 1.0$ GHz of $Q = 68$ at $V_r = -4V$ or $Q = 114$ at $V_r = -10V$ (or $f = 10.0$ GHz values of $Q = 6.8$ and $Q = 11.4$, respectively). Considering

20      that an interesting X-band ($f = 10$ GHz) RADAR application might want a bandwidth of $f = 20$MHz for the full width at half-maximum (FWHM), corresponding to a $Q = 500$ quality factor,

we see that available varactors have inadequate Q (too much loss) to meet such requirements.

While the mechanisms are different, this will very likely apply to the use of ferroelectrics or other

"tunable materials." A general characteristic of materials which exhibit the field-dependent

dielectric constant nonlinearities (that makes them tunable) is that they exhibit substantial values

5    of the imaginary part of the dielectric constant (or equivalently, loss tangent). This makes it

unlikely that, as in varactors, these "tunable materials" will be capable of achieving high Q's,

particularly at high signal frequencies.

[0008]    An additional problem with both varactors and "tunable materials" for

circuits with high values of Q is that these are basically two-terminal devices; that is, the dc tuning

10    voltage must be applied between the same two electrodes to which the signal voltage is applied.

The standard technique is to apply the dc tuning bias through a "bias tee"-like circuit designed to

represent a high reactive impedance to the signal frequency to prevent loss of signal power out the

bias port (as this would effectively reduce the Q). However, while the design of bias circuits that

limit the loss of energy to a percent, or a fraction of a percent, even losses of a fraction of a percent

15    are not nearly good enough for very high Q circuits (*e.g.*, Q's in the $10^3$ to $>10^4$ range, as

achievable with HTS resonators). It would be much easier to design such very high Q circuits

using three-terminal, or preferably 4-terminal (two-port) variable capacitors in which the tuning

voltage is applied to a completely different pair of electrodes from those across which the signal

voltage is applied (with an inherent high degree of isolation between the signal and bias ports).

20    [0009]    One new form of variable capacitor which avoids the problems of varactors

or "tunable materials" approaches is the microelectromechanical (MEMS) variable capacitor such

as that disclosed by U.S. Pat. No. 5,696,662. A number of MEMS variable capacitor device

structures have been proposed, including elaborate lateral-motion interdigitated electrode capacitor

structures. In the simple vertical motion, parallel plate form of this device, a thin layer of

dielectric separating normal metal plates (or a normal metal plate from very heavily doped silicon)

5    is etched out in processing to leave a very narrow gap between the plates. The thin top plate is

suspended on four highly compliant thin beams which terminate on posts (regions under which the

spacer dielectric has not been removed). The device is ordinarily operated in an evacuated

package to allow substantial voltages to be applied across the narrow gap between plates without

air breakdown (and to eliminate air effects on motion of the plate and noise). When a dc tuning

10   voltage is applied between the plates, the small electrostatic attractive force, due to the high

compliance of the support beams, causes substantial deflection of the movable plate toward the

fixed plate or substrate, increasing the capacitance.

[0010]    Because the change of capacitance, at least in the metal-to-metal plate

version of the MEMS variable capacitor, is due entirely to mechanical motion of the plate (as

15   opposed to "instantaneous" electronic motion effects as in varactors or "tunable materials"), the

frequency response is limited by the plate mass to far below signal frequencies of interest.

Consequently, these MEMS devices will be free of measurable intermodulation or harmonic

distortion effects, or other dynamic range problems (up to the point where the combination of bias

plus signal voltage across the narrow gap between plates begins to lead to nonlinear current

20   leakage or breakdown effects).

[0011]     In addition to their freedom from IM/dynamic range problems, normal metal plate MEMS variable capacitor structures offer the potential for substantially lower losses and higher Qs. While the simple parallel plate MEMS structure, which uses the top and bottom plates as the two capacitor electrodes, has a Q problem due to the skin effect resistance, $R_{ac}$, of the long narrow metal leads down the compliant beams supporting the movable plate, an alternative structure is possible which avoids this problem. If the top (movable) plate is made electrically "floating" (from a signal standpoint; it would still have a dc bias lead on it), and the fixed bottom plate split into two equal parts, these two split plates can be used as the signal leads to the MEMS variable capacitor. (The capacitance value is halved, of course, but the tuning range is preserved.) In this "floating plate" configuration, passage of ac current through the long narrow beam leads is avoided, allowing fairly high values of Q to be achieved, even with normal metal plates.

[0012]     While this conventional MEMS variable capacitor structure is capable of improved Qs and avoids the intermodulation problems of varactors and "tunable materials," it has some potential problems of its own. Because only the relatively weak electrostatic attraction between plates is used to drive the plate motion to vary the capacitance, the plate support "spider" structure must be extremely compliant to allow adequate motion with supportable values of bias voltage. A highly compliant suspension of even a small plate mass may render the device subject to microphonics problems (showing up as fluctuations in capacitance induced by mechanical vibrations or environmental noise). Having the electric field which drives the plates directly in the (vacuum) signal dielectric gap may cause another problem. In order to achieve a high tuning range (in this case, the ratio of the capacitance with maximum dc bias applied to that with no dc bias),

the ratio of the minimum plate separation to the zero-bias plate separation must be large (*e.g.*, 10x would be desirable). Unfortunately, the minimum gap between the plates (maximum capacitance, and correspondingly, maximum danger of breakdown or "flash-over" failure between the plates) is achieved under exactly the wrong bias conditions: when the dc bias voltage is at a maximum.

5      [0013]      Therefore, there is a need in the art for a new tunable filters incorporating MEMS-like HTS variable capacitor structure which offers extremely high Q wide tuning range, freedom from microphonics and breakdown problems, and fully isolated port tuning.

## SUMMARY OF THE INVENTION

10      [0014]      In one innovative aspect, the present invention comprises a tunable filter circuit including a variable capacitor. The variable capacitor comprises a fixed substrate on which is disposed a fixed plate(s). Preferably, the fixed plate is comprised of a high-temperature superconductor material. Adjacent to the fixed substrate is a movable substrate on which is a second plate comprised of either a normal metal or (for even higher Q) a high-temperature
15    superconductor material. A filter signal voltage is coupled to the fixed and second plates that are disposed between the fixed and movable substrates and substantially parallel to one another such that the fixed and second plates define a capacitor gap(s) across which the filter signal voltage is impressed. A mechanical driver, operatively coupled to a tuning signal, is mounted on the fixed substrate and mounted to the movable substrate whereby mechanical displacements (for example,
20    length displacements or other changes in shape) in the driver, in response to changes in the tuning signal, causes changes in the magnitude of the gap separating the movable plate and the fixed

plate. Because the fixed and second plates define a capacitor, such gap alterations affect the capacitance(s), and in turn affect the functioning of the tunable filter circuit. Thus, a change in the tuning signal alters the frequency response of the tunable filter.

[0015]     In a preferred embodiment, the fixed plate is split so that a first fixed plate

5      and a second fixed plate both comprising a high-temperature superconductor material are on the fixed substrate. A floating plate, also comprising a high-temperature superconductor material is on the movable substrate. The fixed plates and the floating plates are disposed between the fixed and movable substrates so as to define a capacitor gap. Rather than coupled across the capacitor gap, the filter signal is coupled to the first and second fixed plates. Alternatively, the first and

10     second fixed plates and the floating plate may comprise a normal metal (such as Ag) although such a substitution lowers the achievable Q.

[0016]     The mechanical driver provides a robust, fast and well-controlled variation in the gaps(s) between the movable plate and the fixed plate(s) of the variable capacitor. The mechanical driver may provide relatively high forces over plate gap changes of at least a few

15     microns (and preferably 10 microns). The mechanical driver may incorporate conventional piezoelectric or electrostrictive ceramic actuator materials such as PZT (lead zirconate titanate) which provide length displacements on the order of $10^{-3}$ (the PZT actuator changes its total length approximately 0.1% in response the tuning signal). Alternatively, the mechanical driver may incorporate newer classes of materials such as higher-performance piezoelectric ceramics or

20     electroactive polymer materials which provide a greater achievable mechanical displacement with respect to the dimensions of the mechanical driver. In response to a tuning signal, such materials

have mechanical displacements that are more complex than the simple length displacements induced in a PZT actuator. However, the geometry of the mechanical driver may be altered to allow these alternate materials to be utilized.

[0017]     In a preferred embodiment wherein the mechanical driver incorporates a length-displacement actuator material such as PZT, the mechanical driver comprises a folded piezoelectric or electrostrictive structure coupled to a tuning signal wherein a first piezoelectric driver is attached to the fixed substrate and to a coupling member. A second piezoelectric driver is also attached to the coupling member and to the movable substrate wherein a length decrease in the first piezoelectric driver and a length increase in the second piezoelectric driver causes the capacitor gap to decrease. The tunable filter may further comprise an HTS inductor coupled between the first and second fixed plates. Preferably, the HTS inductor has an electrical length chosen to support resonance either at the HTS inductor's fundamental resonant frequency or at harmonics of the fundamental resonant frequency.

[0018]     The present invention avoids intermodulation problems by the use of a mechanical driver to affect the capacitance because the masses involved in the mechanical tuning processes do not allow substantial motion of the capacitor plates to occur at the signal frequencies of interest. In addition, the present invention permits the tuning signal and the filter signal to be de-coupled, avoiding the prior art problems associated with two-terminal designs. The present invention offers an extremely high Q, driven in one embodiment by piezoelectric actuators for tuning high speed and low microphonics.

## DESCRIPTION OF THE DRAWINGS

[0019]     Fig 1a is a cross section of a variable capacitor according to one embodiment of the invention.

5     [0020]     Fig 1b is a plan view, partially cut away, of a variable capacitor structure incorporated in a resonator forming a tunable filter according to one embodiment of the invention.

[0021]     Fig. 1c illustrates an equivalent circuit of the filter resonator structure of Fig. 1b.

[0022]     Fig 1d is a cross section of a variable capacitor according to one

10     embodiment of the invention.

[0023]     Fig 1e is a cross section of a variable capacitor according to one embodiment of the invention.

[0024]     Fig. 2a is a plan view of a one-half wavelength resonator circuit.

[0025]     Fig. 2b is a plan view of an equivalent circuit representation of the half-

15     wavelength resonator circuit illustrated in Fig. 2a.

[0026]     Fig. 2c is a plan view of a one-half wavelength resonator circuit.

[0027]     Fig. 2d is a plan view of an equivalent circuit representation of the half-wavelength resonator circuit illustrated in Fig. 2c.

[0028]     Fig 3a is a cross section of a variable capacitor according to one embodiment of the invention.

[0029]     Fig 3b is a plan view, partially cut away, of a variable capacitor structure according to one embodiment of the invention.

5     [0030]     Fig. 4 is a graph relating the insertion loss vs. frequency relationship in a loop-coupled single-resonator tunable filter according to one embodiment of the invention.

[0031]     Fig. 5 is a graph relating the insertion loss vs. frequency in a loop-coupled single-resonator tunable filter according to one embodiment of the invention.

[0032]     Fig. 6a is a cross section of an electrostatic-actuated capacitor in its low

10     capacitance state.

[0033]     Fig. 6b is a cross section of an electrostatic-actuated capacitor in its high capacitance state.

[0034]     Fig. 6c is plan view of a digitally tunable HTS filter.

15     <u>DETAILED DESCRIPTION OF THE INVENTION</u>

[0035]     Turning now to Figure 1a and 1b, a preferred embodiment for a high-Q bandpass filter resonator based on a MEMS-like HTS split-plate variable capacitor with independent mechanical actuation is illustrated. A first fixed plate 10 and a second fixed plate 15 are patterned from thin-film HTS material (*e.g.,* YBCO) grown on a fixed substrate 20 (see, *e.g.,*

Figure 1a) comprising, for example, MgO. Such epitaxial superconductive thin films are now

routinely formed and commercially available. See, *e.g.*, R. B. Hammond, et al., "Epitaxial

$Tl_2Ca_1Ba_2Cu_2O_8$ Thin Films With Low 9.6 GHz Surface Resistance at High Power and Above

77K," Appl. Phy. Lett., Vol. 57, pp. 825-27, 1990. The variable capacitor structure 25 is

5    completed by the addition of a floating plate 30 (drawn transparent in the plan view in Figure 1b)

patterned from thin-film HTS material grown on a movable substrate 35 (see, *e.g.*, Figure 1a)

comprising, for example, MgO. Floating plate 30 is spaced apart and substantially parallel to fixed

plates 10 and 15 and may completely cover the fixed plates 10 and 15. As a result, the HTS

variable capacitor structure 25 actually comprises two variable capacitors in series, (the equivalent

10   circuit is illustrated in Fig. 1c) which halves the capacitance per unit area over that of a normal

parallel plate structure. The advantage is that no conductive contact to floating plate 30 is

required, which greatly simplifies (particularly for an HTS implementation) the achievement of

very low series resistance contact to the capacitor (for high Q). In such an embodiment, a filter

signal is coupled only to the fixed plates 10 and 15. A mechanical driver 45 (see, *e.g.*, Figure 1a),

15   which in one embodiment is coupled between the fixed substrate 20 and the movable substrate 35,

adjusts the capacitor gap 50 (see, *e.g.*, Figure 1a), preferably maintained in a vacuum, defined by

the fixed plates 10 and 15 and the floating plate 30. Those of ordinary skill in the art will

appreciate that a number of alternative mechanical drivers are suitable, *e.g.*, piezoelectric drivers

operating in various modes or drivers comprising electroactive polymer materials, provided that

20   suitable modifications to the mechanical driver geometry are incorporated.

[0036]     In this floating plate HTS variable capacitor tunable filter structure 25, wide

HTS lines (potentially much wider than illustrated in the top view of Figure 1b) may be used to

define an inductor 40 (or alternatively, a transmission line segment) coupled at full width between

the fixed plates 10 and 15 to achieve very high Q values.  The inductor 40 may have an electrical

5      length substantially equal to a one-half wavelength corresponding to a fundamental resonant

frequency so that it becomes a simple L-C resonator circuit.  Alternatively, the inductor 40 may

have an electrical length substantially equal to multiples of this one-half wavelength.  Moreover,

the inductor 40 may be substantially shorter than one-half wavelength, thus approximating a

lumped circuit element.  The resulting equivalent circuit is portrayed in Fig 1c.  Two variable

10     capacitors 60 and 65 are in series with the inductor 40.  The mechanical driver 45 varies the

capacitance of capacitors 60 and 65.

[0037]     The resulting equivalent circuit as portrayed in Fig 1c has important

consequences resulting from its balanced structure.  Note that a commonly used half-wavelength

resonator comprises an open circuited length of microstrip transmission line as illustrated in Figure

15     2a.  The transmission line has a length (denoted as "d") substantially equal to one-half wavelength

corresponding to the fundamental resonant frequency $f_0$.  Such resonators are frequently tuned by

use of a metal tuning screw through the cover plate near one end of the resonator to add

capacitance between the end of the resonator and ground.  Instead of a screw, one could use a

variable capacitor $C_t$ as disclosed by the present invention.  For frequencies at or near resonance,

20     the impedance $Z_{in}$ seen looking into the end of the resonator matches that of a parallel tuned circuit

as illustrated in 2b. For a given capacitance variation $\Delta C_t$, there will be some resulting resonant

frequency variation $\Delta f_0$.

[0038]    Should the half-wavelength ($\frac{\lambda_0}{2}$) resonator of Figure 2a be bent into a

"hairpin" configuration as shown by the inductor 40 in Figure 1b and $C_t$ be connected across the

5    open ends of the resonator, the balanced structure of Figure 2c is produced. At resonance, the

potentials on the two open ends of this "hairpin" resonator have opposite potential, so that the

terminals of the variable capacitor are now driven in a balanced manner. Because of the balanced

operation, for whatever potential that exists at a point on the right side of the hairpin, there is an

equal and opposite potential at the corresponding point on the left side of the resonator. Therefore,

10    a virtual ground will run through the centerline of the resonator as shown in Figure 2d. It is then

convenient to model the capacitance in the equivalent form wherein the single capacitor $C_t$ is

replaced by two capacitors each having twice the capacitance of the original capacitor $C_t$. If the

circuit illustrated in Figure 2d were straightened out as in Figure 2a, there would be a capacitor of

size 2Ct at both ends of the transmission line. Thus, the capacitor Ct in Figure 2c is four times

15    more effective for tuning than it is in the structure illustrated in Figure 2a.

[0039]    The present invention provides a useful tunable filter possessing an

extremely high Q value. The following example illustrates operation of the present invention at

9.70 GHz. Those of ordinary skill in the art will appreciate that such an example is merely

illustrative of the invention and that the example is easily scaled to illustrate operation at lower or

20    higher frequencies. As an illustration, consider a simple X-band L-C resonator of the type shown

in Figure 1a, having C = 5.0 pF (actually two 10 pF capacitors in series, wherein the first and

second fixed plates 10 and 15 are 0.75mm square if the plate spacing is 5.0 μm). To give a

resonant frequency of f = 9.70 GHz requires an inductor 40 having a value of L = 0.5384 nH,

which would be produced by a loop having a nominal diameter of D = 2.0 mm if its inside

5      diameter is 1.0 mm. This results in an inductor 40 having an outside diameter of 3.0 mm and an

inner hole diameter of 1.0 mm). Because the width (1 mm) of the HTS conductor forming the

inductor loop 40 is nearly the same as the dimensions of the first and second fixed capacitor plates

10 and 15, the capacitor and inductor structures can be smoothly mated with little loss of Q.

While, due to current crowding at the conductor edges (particularly at the edges of the hole), it is

10     difficult to calculate the exact ac resistance, if we assume an operating frequency f = 10 GHz and

an HTS surface resistance of $R_S$ = 0.1 mΩ (1 x $10^{-4}$ ; typical for high quality YBCO at 80° K), and

then account for a 2x current crowding factor by only considering contribution from one of the two

conducting HTS surfaces, the ac resistance $R_{ac}$ is given by $R_S$ (length/width) = $R_S$ (πD) (w)

=0.628mΩ for D = 2mm and w = 1 mm. Therefore, at an operating frequency of 10 GHz, the

15     calculated unloaded Q of this 2mm diameter HTS loop inductor 40 would be Q = 2 πfL/ $R_{ac}$ =

53,840, a Q value far higher than most designs of interest require. Assuming that the loss tangent

of the MgO HTS growth substrates 20 and 35 is sufficiently low and energy losses through

radiation or coupling to other structures are minimal, loaded Q values of 1000 or more should be

achievable, even in simple single-resonator filters having very low insertion loss.

20     [0040]      A number of alternate geometries may be employed to construct the variable

capacitor used in the present invention. For example, rather than employing the preferred "split

plate" design illustrated in Figure 1a, the first and second fixed plates 10 and 15 could be replaced

with a single fixed plate (not illustrated). In such a design, however, both plates on either side of

the capacitor gap 50 would require a coupling to the filter signal. In addition to alternate

geometries, the floating plate 30 or the first and second fixed plates 10 and 15 may be alternatively

5      constructed out of gold or a similar low-loss metal rather than from HTS material. Such metals do

not produce, however, Q values as high as that can be achieved using an all HTS construction.

[0041]      Turning now to Figure 1d, an embodiment of the invention is illustrated in

which the movable substrate is sandwiched between two fixed substrates 20 and 21. The first and

second fixed plates 10 and 15 are on the fixed substrate 20 and the floating plate 30 is on the

10     movable substrate 35 as previously described. However, mechanical driver(s) 45 are coupled

between the fixed substrate 21 and the movable substrate 35 so that the mechanical driver(s) 45 do

not span across the capacitor gap 50 as did the mechanical driver 45 in the embodiment illustrated

in Fig 1a. Alternatively, turning now to Figure 1e, the first and second "fixed" plates 10 and 15

could also be on a movable substrate 35 positioned by additional mechanical driver(s) 45

15     analogous to the embodiment illustrated in Fig 1d.

[0042]      While the performance potential for the HTS variable capacitor tunable

filters within the scope of the present invention is very exciting, the fabrication of such devices

will now be addressed; specifically, how to achieve the required precise motion of the floating

plate 30. A fast, stiff, robust mechanical driver 45 is very desirable. An appealing mechanical

20     driver approach, because of its simplicity and low cost, is to use conventional piezoelectric

actuators. Figures 3a and 3b illustrate one embodiment of the invention for implementing an HTS

variable capacitor tunable filter structure 25 (see Figure 3a) with piezoelectric drivers. A unique

design feature in Figures 3a and 3b is the use of a "folded" piezoelectric driver 70 structure on

each end of the movable substrate 35 (best seen in Figure 3a). The reason for this unusual

structure is that, while the stresses induced in piezoelectric materials by the application of electric

5      fields are large, their elastic modulus is also high, so the actual attainable "dc" elongations are only

of the order of $10^{-3}$ times the length of the material (higher "ac" elongations are achieved at

resonance, but that is not usable for this variable capacitor application). For example, if we chose

a nominal plate separation of 5 µm, variable from near zero to 10 µm with a ±5 µm range of

motion, then we need at least a 5 millimeter length of piezoelectric material. By "folding" this

10     piezoelectric material length into 2 pieces (one going up and one going down), we can keep the

total height under 3 mm, and automatically cancel the effects resulting from the substantial CTE

(coefficient of thermal expansion) of the piezoelectric material.

[0043]        The basic structure of this embodiment of the variable capacitor HTS

tunable filter 25 is similar to that already discussed with respect to Figs. 1a and 1b. As best seen in

15     Figure 3a, to accommodate the two folded piezoelectric driver structures 70 and to minimize

parasitic capacitive effects between the variable capacitor HTS tunable filter 25 and the

piezoelectric driver structures 70, the movable substrate has lateral extensions 80 and 85 which

extend laterally substantially past fixed plates 10 and 15, respectively. The first fixed plate 10 and

the second plate 15 are patterned from HTS material and grown on the fixed substrate 20

20     comprising, for example, MgO. An HTS material inductor 40 (not illustrated) is coupled between

the first and second fixed plates 10 and 15. As previously described, this inductor 40 preferably

has an electrical length chosen such that the inductor 40 behaves as a simple L-C resonator circuit.

As illustrated in Fig 3b, the first and second fixed plates 10 and 15 each has a lead 16 so that a

filter signal may be coupled to the variable capacitor HTS tunable filter 25 (seen in Figure 3a).

The variable capacitor HTS tunable filter structure 25 is completed by the addition of a floating

5    plate 30 (drawn transparent in the plan view in Figure 3b) patterned from thin-film HTS material

grown on the movable substrate 35 comprising, for example, MgO. As seen in Figure 3a, a pair of

folded piezoelectric drivers 70 is coupled to the lateral extensions 80 and 85 as follows. A first

piezoelectric driver 100 is attached, using for example, low-temperature indium alloy solder

(which also may be used for the remaining required piezoelectric attachments), to the fixed

10   substrate 20 at a location adjacent to the lateral extension 80 of movable substrate 35. The first

piezoelectric driver 100 is attached at its opposite end to a coupling member 110. Also attached to

coupling member 110 is the second piezoelectric driver 115. Coupling member 110 may be

constructed from a suitable material such as MgO. Alternatively, coupling member 110 may

simply be a low temperature indium solder bond or the like in order to couple the first and second

15   piezoelectric drivers 100 and 115 together. The second piezoelectric driver 115 is attached at its

opposite end to the lateral extension 80 of movable substrate 35, forming one of the pair of folded

piezoelectric driver structures 70 wherein the first piezoelectric driver 100 and the second

piezoelectric driver 115 are substantially parallel to one another. In this fashion, when the first

piezoelectric driver 100 shortens and the second piezoelectric driver 115 lengthens, the gap 50

20   between the capacitor plates is shortened, affecting the tunable filter 25 frequency response. The

other folded piezoelectric driver structure 70 is coupled to the lateral extension 85 of movable

substrate 35 in an analogous fashion as follows. A third piezoelectric driver 120 is attached to

lateral extension 85 at one end and attached to a coupling member 125 at its opposite end. Also attached to coupling member 125 is a fourth piezoelectric driver 130. The fourth piezoelectric driver 130 is attached at its opposite end to the fixed substrate 20 at a location adjacent to lateral extension 85 so that the third piezoelectric driver 120 and the fourth piezoelectric driver 130 are

5    substantially parallel to one another, thereby forming the other piezoelectric driver structure 70. To suppress bending vibrations, an optional bridge 140, comprised of, *e.g.*, MgO, may span between the coupling members 110 and 120.

[0044]    In one embodiment, each piezoelectric driver is coupled to a tuning signal having a tuning signal voltage, Vt through electrodes 105. To implement the folded piezoelectric

10    actuator structure 70, a proper combination of tuning signal voltage (Vt) polarity and the orientation of the piezoelectric material ("poling" direction) is necessary. For example, in one embodiment, with positive voltage Vt, the first and fourth piezoelectric drivers 100 and 130 get shorter and the second and third piezoelectric drivers 115 and 120 lengthen, driving the floating plate 30 closer toward the fixed plates 10 and 15, decreasing the gap 50 and increasing the

15    capacitance. Analogously, in the same embodiment, a negative voltage Vt would lengthen the first and fourth piezoelectric drivers 100 and 130 and shorten the second and third piezoelectric drivers 115 and 120. Thus, with a negative voltage Vt, the gap 50 between the plates increases, decreasing the capacitance. The mode of piezoelectric material (typically lead zirconate titanate, "PZT") operation is as illustrated at the lower left in Figure 3a, in which a bias voltage, Vt, applied

20    across the thin "Z" dimension of a plate causes either shortening in "X" and lengthening in "Y," or lengthening in "X" and shortening in "Y," depending on the polarity of voltage Vt. In such an

embodiment, the orientation of the paired piezoelectric drivers with their respective electrodes

within each folded piezoelectric driver structure may be denoted as "opposite" to each other.

[0045]     Roughly analogous "opposite" paired piezoelectric driver structure

(including the same piezoelectric mode and polarities) is used in standard commercial "bimorph"

5     beam actuators.  A principal difference is that in a standard bimorph beam, a thin metal (typically

5mil stainless steel) shim fills the gap between the two PZT layers (typically 7.5mils thick), such

that when bias is applied and the PZT on one side lengthens and the PZT on the other side

shortens, the beam bends.  In the present invention, mechanically decoupling the two PZT drivers

(except at the end attached to the coupling members 110 and 125), and preventing bending of the

10     PZT material (by the symmetrical structure, optionally augmented by putting an MgO bridge 140

across the coupling members 110 and 125 as shown in Figure 3a) leads to a Vt bias dependent

change in height between the bottom ends of the piezoelectric drivers.  For those not familiar with

piezoelectric actuator materials and structures, the paper "Modeling of a Three-Layer Piezoelectric

Bimorph Beam with Hysteresis" by T. S. Low and W. Guo, published in the Journal of

15     Micromechanical Systems, Volume 4, No. 4 [December 1995], pages 230-237 would be a useful

reference.

[0046]     The HTS variable capacitor structure requires, at least in the vertical

dimension, precision mechanical fabrication and assembly.  The precision fabrication is expressed

principally as a flatness and peak non-planarity (bump height) specification on the facing surfaces

20     of the fixed plates 10 and 15 and floating plate 30.  To avoid short circuiting between opposing

capacitor plates as the mean separation distance between the surfaces is reduced to the micron

range (or even submicron range), there must not be any substantial protrusions from the facing

surfaces of the plates. In one embodiment, the as-grown HTS surfaces are scanned for the

presence of nodules or other "bumps" protruding from the surface and laser ablation or other

techniques are employed to remove any that are too high. Note that in a capacitor plate, there is no

5      significant performance penalty for having a few holes in the conductor film, so such a post-

growth "re-figuring" of the surfaces need not be planarizing, just effective in removing "bumps"

("dips" are not a problem). The critical problem in precision assembly is establishing the correct

initial (Vt = 0) height (separation) and parallelism of the capacitor plates, preferably to a fraction

of a micron (at least in parallelism). In one embodiment, this precision vertical alignment may be

10     achieved by means of a self-alignment process based on an art-recognized type of removable

deposited spacer technique. For example, if a 5 μm nominal spacing were desired, then a uniform

5 μm (post-develop) thick layer of photoresist could be spun onto the previously-patterned HTS

fixed substrate 20 (shown as MgO in Figure 2) and exposed and developed to leave, for example,

six "pucks," 5μm high (one at each corner and two along the long sides of where the movable

15     substrate 35 will be placed. The movable substrate 35 may then be aligned and pressed down onto

these posts while the piezoelectric drivers 70 are attached on each end with low-temperature (*e.g.*,

indium alloy) solder. After assembly is complete, the posts are dissolved away in a suitable

solvent to free the floating plate 30 for unimpeded vertical motion when tuning bias voltage, Vt, is

applied. (Note that the pucks may be made of photosensitive material, or other, more easily

20     dissolved material (like PMMA) which has been patterned using photoresist.)

[0047]     Note in Figure 3a that the thinned (for low mass when fastest tuning rate is desired) movable substrate 35 extends considerably beyond the ends of the HTS floating plate 30 itself, forming lateral extensions 80 and 85.  The reason for these substantial gaps between the floating plate 30 and the attachment points for the second and third piezoelectric drivers 115 and

5     120 is to minimize stray capacitance between the HTS floating plate 30 and the folded PZT driver structures 70 to avoid energy loss/Q degradation effects.  Note that this piezoelectric actuator approach provides the ideal of a tuning input (the Vt electrodes on the PZT drivers) that is completely separate from the signal path; leaving a good clearance between the floating plate 30 and the PZT drivers just insures that maximum Q is realized.

10     [0048]     To better illustrate what even a very simple tunable HTS filter of the type illustrated in Figures 1a, 1b, 3a, and 3b could do in a system application, an example of a simple, single-resonator, narrowband bandpass preselector filter for an X-band RADAR was considered. Those of ordinary skill in the art will appreciate that such an example is merely illustrative of the invention and that the example is easily scaled to illustrate operation at lower or higher

15     frequencies.  The $L_2 = 0.5384nH$, $C_2 = 0.50$ pF (for $f_0=9.70GHz$) parallel "tank" resonator (the dimensions and projected unloaded Q of which were discussed above) was considered with simple inductive loop coupling to a 50 ohm source and a 50 ohm load was considered.  Analysis of the circuit was performed both analytically and using simulated program with integrated circuit emphasis (SPICE) (the two approaches give the same result).  Since the unloaded Q of the

20     resonator (approximately 50,000 as calculated previously) is so much higher than the intended loaded Q (a bandwidth of 20MHz [full width at half maximum (FWHM)] at $f_0 = 10$ GHz would be

appropriate for a RADAR application, corresponding to a loaded $Q = f_0/ \Delta f = 500$), the ac

resistance $R_{ac}$ of the $L_2$-$C_2$ resonator was ignored. The bandwidth of this simple loop-coupled

single resonator tunable bandpass filter is determined by the inductance of the input and output

coupling loops ($L_1$ and $L_3$, both taken as 20% of $L_2$) and their coupling coefficients ($K_{12}$ and $K_{23}$,

5     both taken as 0.09) to the resonator inductor, $L_2$. These particular values were selected to give a -

3dB bandwidth of 20MHz. Note that in the physical construction of such an HTS filter, it is not

necessary that these input and output coupling loops be made of HTS material, or even lie on the

cooled HTS substrate. It would be easier to have them lie well above and below the resonator

substrate (to minimize direct coupling between them, and to have their positions mechanically

10     adjustable when the filter is setup to select the exact bandwidth desired. Note that placing these

input and output loops out of contact with the cooled HTS substrate avoids the thermal conduction

load associated with metallic conductors to the substrate. (This advantage would apply equally

well to capacitive probe coupling to the resonator, which is another satisfactory means of making a

resonator into a bandpass filter).

15     [0049]     Figure 4 shows a series of insertion loss vs. frequency curves over a 0 to -

35dB insertion loss and a 9.40GHz to 10.0 GHz frequency range. The inductance values and

coupling coefficients are the same for all curves; only the value of the variable capacitor, $C_2$, is

changed. The curves are plotted for nominal (unloaded) $L_2$-$C_2$ resonant frequencies of 9.40GHz,

9.45GHz, 9.50GHz ... to 10.0 GHz. ($L_2 = 0.538427$ nH for all curves, with $C_2$ calculated from the

20     equation $C_2 = (2 \pi f_0)^2 L_2$. The same data is plotted in more detail over a 0 to -15 dB insertion loss

range over a more limited 9.59 GHz to 9.81 GHz frequency range (10 MHz per major division) in

Figure 5. We see in Figure 5 that the -3dB bandwidth is indeed 20 MHz (f $_{-3db}$ = f$_0$ ± 10 MHz) at

f$_0$= 9.7GHz. While the bandwidth changes as the center frequency, f$_0$, is changed (increasing or

decreasing in proportion to f$_0$) the change over the X-band frequencies of interest (Figure 4) is not

very large. Given a 10:1 plate separation range (*e.g.*, 1 µm to 10 µm), the achievable capacitance

tuning range (including fringe capacitance) should be at least 9:1, which would imply a 3:1

frequency tuning range. The insertion loss of the filter is negligible, due to the high unloaded Q of

the HTS tunable resonator. Of course, with only a single resonator, for a given bandwidth, the

rejection outside of this passband is limited to –20 dB to –40 dB (depending on frequency

difference from passband center frequency to interferer). Considering that the purpose of a

RADAR preselection filter is not to define the bandwidth, just to push down the amplitude of

interfering signals or jammers in order to reduce intermodulation or de-sensitization effects on

system performance, this level of out of band rejection would be fine. Of course, much higher out

of band performance (as well as flatter in-band response) could be achieved by making use of

multiple tunable resonator assemblies coupled together (using standard filter design technology) in

a tunable filter having many more poles.

[0050]     From an RF/µW performance standpoint (unloaded Q, tuning range, etc.)

the present invention offers outstanding performance. There is some question of whether the

tuning speed, using only conventional PZT materials, can be brought down to the 5 µs to 10 µs

range that would be desirable for frequency-hopping applications (unless operation with very small

gaps proves feasible, although the use of higher performance ceramic piezoelectric or electroactive

polymer materials should make this possible). However, the filters themselves are quite small, and

so there would be little penalty to putting two of these filters in a cryocooler vacuum dewar and

"ping-ponging" back and forth between them on successive hops. Thus, with a hopping residence

time of a few milliseconds, there would be plenty of time to tune the "off duty" filter to the next

hop frequency. In fact, even if undesirable piezoelectric driver hysteresis or drift characteristics

5     prevent obtaining precise "first-shot" open-loop tuning, there should be ample time to check the

new set frequency of the "off–duty" filter and correct it if necessary. Of course, in a preferred

embodiment, the present invention will use the finest available "AFM quality" piezoelectric

actuator technology to achieve fast, open-loop operation in the simplest way possible.

[0051]        While Figs 3a and 3b illustrate a piezoelectrically-actuated form of the

10    microelectromechanical (MEMS) continuously-variable capacitor structure, the present invention

includes other means of actuating a variable capacitor structure. For example, Figure 6a and 6b

illustrate an electrostatically-actuated form of MEMS capacitor structure capable of switching

between a high and a low capacitance state. The switched capacitor is formed on a substrate 150

(comprised of, for example, quartz) upon which is deposited a metal electrode 155. The electrode

15    155 is sandwiched between the substrate 150 and a dielectric layer 160. A flexible metal

membrane 170 is formed or biased so as to be spaced apart from the dielectric layer 160, thus

forming a low capacitance state as illustrated in Figure 6a. When a suitable electrostatic potential

is applied across the electrode 155 and membrane 170, the flexible membrane 170 deflects against

the dielectric layer, inducing a high capacitance state as illustrated in Figure 6b. Preferably, both

20    the flexible metal membrane 170 and the electrode 155 are formed from gold or a suitable low loss

metal and are cryogenically cooled during operation. During such cryogenic operation, the

switched capacitor is expected to have a Q of 20,000 at a frequency of 1 GHz.

[0052]     Consider an array of N such switched capacitors arranged in parallel,

wherein N is an integer. The capacitors could be formed so that a first capacitor would have a

5     given capacitance value, a second capacitor would have twice the capacitance value of the first and

so on. A control circuit would determine the amount of capacitance required to, for example, tune

a tunable HTS filter and generate a digital command signal corresponding to the required

capacitance. A decoder would switch one or more of the switched capacitors in response to the

digital command signal. Thus, the array would form a binary sequence with the first capacitor

10     corresponding to bit one of the digital command signal, the second capacitor corresponding to bit

two of the digital command signal and so on. Such a capacitor array, showing bit one, bit two, and

bit three, is illustrated in Figure 6c. An HTS resonator 40 would be coupled across the array in

order to complete the tunable filter. The resonator 40 may have an electrical length substantially

equal to a half-wavelength corresponding to the fundamental resonant frequency of the resonator

15     40 or it may have an electrical length substantially equal to multiples of this half-wavelength.

Tuning is accomplished in discrete steps (digital tuning) with a granularity that depends on the

number of bits in the sequence.

[0053]     The above detailed description and the accompanying drawings are intended

to illustrate specific embodiments of the present invention and are not intended to limit the scope

20     of the present invention. Those of ordinary skill in the art will recognize that various

modifications of the above-described embodiments are possible without departing from the spirit

of the invention. Therefore, the scope of the present invention is defined by the following claims.